55

COHERENT RECEPTION USING CARRIER LOCK AND SIDEBAND LOCK TECHNIQUES*

PART I

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ABSTRACT. In this paper the design, construction and performances of a receiving system using coherent reception techniques, have been described. The principles of Carrier Lock and Sideband Lock techniques have been discussed. It is pointed out that the reception of different types of modulations (*viz.*, DSB, SSB, conventional AM, Narrowband PM or FM, Binary PSK and VSB) can be made utilising CPL and SBPL techniques. The effects of noise and interference on the locking capability for the system have been studied. The loop phase equations are given in both the cases and modifications that will be caused in presence of noise are also pointed out. This is followed by a description of specific circuits necessary for implementation of the techniques of reception of different modulations mentioned. Experimental results are given with regard to performances of the receiver in the presence of noise, interference and carrier jitter; the minimum bandwidth necessary for the controlling loop for successful tracking of the modulation envelope. Suggestions are given to extend this technique further in reception of narrowband FSK signal.

INTRODUCTION

The transmission of a message is effected by varying some characteristics of radio frequency wave in accordance with the message or the intelligence to be transmitted. The reception of the signal consists in the recovery of the transmitted message from the received modulated signal i.e. from the received carrier and sideband components in general. The basic idea in reception is to correlate the message or information bearing signal in the received modulated signal and to decorrelate the noise and interference present in the reception band.

A technique of reception is called 'coherent' if specific use is made of the apriori information of all the characteristics of the signals that are not varied in the modulation process for optimum use of the signal energy. In DSB, for example, since the amplitude only of the transmitted signal is modulated, the frequency and the phase of the carrier can be assumed to be constant, the information about these can be stored and used in demodulating (Costas, J. P., 1956). Carrier in the

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receiver should have the same frequency and phase as those of the transmitted signal. Similarly, in reception of M-ary binary keyed signals, use can be made of the information of the expected keying instants and the specific signalling waveforms. Coherent reception is optimum because the fullest possible use is made of the signal energy. It will be realised that coherent reception making use of 'a priori' information stored in the receiver is not practical because the propagation medium is time varying and consequently the characteristics of the transmitted signal even though not modulated, will be subject to unpredictable variations. A practical coherent receiver has therefore to gather these informaltion from the received signal itself, constantly check the stored values with the current information and utilise these for demodulation. The decision process in the demodulator thus involves two steps the first step consists of the 'estimation of the parameters' and the second of 'actual detection' It is important to realise that such a technique is useful only if the time rate of variation of the 'invariant' parameter is relatively small compared to that of the modulated parameter In such a case smoothing and filtering of the data is possible enabling reliable extraction of the parameter concerned

Usefulness of a coherent detection technique arises from the fact that explicit use of the knowledge of the parameters enables one to reject noises and interference which do not have the same character as the signal has.

It would appear that since a coherent receiver is to obtain the estimates of the parameters from the received signal itself, such a receiver is essentially a feedback device. The error signal in the feedback device is a measure of the difference between the present values of the parameters and their past values stored in the system. One can classify the devices in respect of their storage times as (i) feedback detectors if the bandwidth of the control loop (B_c) is small compared to the bandwidth of the message (B_m) and (ii) feedback demodulators if B_c is comparable with B_m It should be obvious that feedback demodulators are essentially tracking devices incorporating a demodulate-remodulate technique and are therefore useful if the time bandwidth product of the message space is large.

The detectors used in coherent receivers are (or intended to be) necessarily linear. A direct consequence is that there is no threshold effect or suppression of the signal by noise or vice-versa; any reduction of interference and noise results from linear filtering alone. In fact, the principal virtue of a coherent reception technique is that it enables, through exploitation of the distinct characteristics of the signal, reception of relatively weak signals in the presence of strong interference.

The block representation of a Coherent or Homodyne (as it is often called) receiver is shown in fig. (1). The ideal reception condition which is necessary for the operation of such a receiver cannot be achieved in practice, because this requires the incorporation of an oscillator in the receiver which is to be in exact frequency and phase synchronism with that in the transmitter all the time during which communication is being made. Even with the use of a crystal controlled

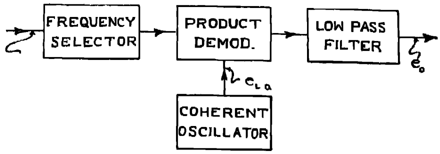

Fig. 1. Block diagram of an ideal coherent reception system.

oscillator with the best engineering attention being given to make their frequencies of oscillation identical, it is not possible to keep their drifts identical. Even if the frequencies of oscillation can be made identical nothing can be reasoned in support of the assumption of the identity in phase of the two oscillations Apart from the question of frequency and phase synchronism, the time varying character of the propagation media cause received signals to fade randomly in time, both in amplitude and phase and in some cases the received signals undergo large 'Doppler shifts'. These facts introduce further complications and thus increase the unpredictability mentioned earlier.

In a practical version implementing this philosophy of reception, the demodulating carrier is brought in time coherence with the modulated one from the received signal itself. The phenomenon of establishing coherence of the local oscillator is called 'Locking' of the demodulating carrier. For satisfactory reception it is required for the local oscillator to automatically lock in phase with the received signal in presence of additive noise, channel perturbation and 'Doppler shifts'. So what is necessary in the receiver is to incorporate specific circuits to establish coherence and to maintain the synchronism for the period during which transmission and reception of the signal are being effected.

A particularly effective method for doing this is to extract the necessary phase information from a pure carrier component and use this as the input to a phase locked loop of narrow effective bandwidth that controls the local oscillator in the receiving system and the mechanism of locking may accordingly be called 'Carrier Phase Lock' technique in coherent reception.

In the reception of suppressed carrier, Binary PSK, Narrowband PM and FM signals, the received sideband components can also be so utilised as to derive the necessary information for establishing and maintenance of the synchronism for the demodulating carrier and the mechanism may accordingly be called 'Sideband Phase Lock' technique in coherent reception. Section that immediately follows deals with the principles of the carrier phase lock (CPL) and sideband phase lock (SBPL) techniques.

Section 3 is devoted to a discussion on locking techniques in reception of different types of modulated signals.

The deleterious effects of noise and interference on the locking capability of the phase locked loop have been dealt with in section 4, which also includes a discussion on the deterioration of the SNR due to noise present in the phase locking loop and derivations of loop-phase equations in presence of noise for (i) carrier phase lock and (ii) sideband phase lock cases. The effect of carrier jitter on the demodulated products has also been considered.

In section 6, the basic circuit necessary and the experimental arrangements for studying the applicability of carrier and sideband phase lock techniques in reception of various types of modulated signals are given. The experimental observations with regard to reception of modulated signals like DSB, Binary PSK, Narrowband PM etc. and the results obtained experimentally with DSB signal corrupted with noise and interference are presented in this section. Some remarks in designing a phase locking loop are also included in this section.

PRINCIPLES OF CARRIER AND SIDEBAND PHASE LOCKING

In this section we shall discuss the carrier-lock and sideband-lock techniques, the philosophy underlying them and the method of implementation of these techniques in practice. The expressions for the controlling d.c. voltage in a system using a SBPL technique will be derived in case of (i) single tone and (ii) multitone modulations.

Carrier-Lock : As mentioned earlier a locked receiver may derive the frequency and phase information from the received modulated wave and make use of these information for demodulating it. A modulated wave consists generally of the carrier and sideband components. The carrier component contains the frequency and phase information and so do the sideband components. In a carrier-lock device the instantaneous phase difference between the received carrier and the local carrier is measured by means of a so-called phase discriminator, the output of the phase discriminator is then used to control the instantaneous phase of the local oscillator in such a way as to reduce and, if possible, eliminate this difference. A reference to fig. (2) will show that the phase equation of phase locking loop is (Chakrabarti, *et al.*, 1964)

$$\frac{d\phi}{dt} = \Omega - Kf(p) \sin \phi \tag{1}$$

where ϕ is the phase difference between the incoming and the local carriers, $f(p)$ is the transfer function of the filter following the discriminator and the constant K is proportional to the amplitudes of the two carriers and the sensitivities of the discriminator and the reactance tube and Ω is the initial difference in frequency between the two carriers.

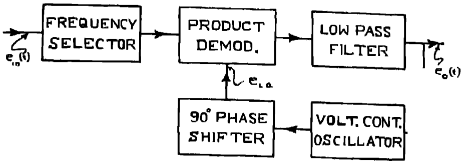

Fig. 2. Block diagram of a carrier phase locked reception system

It follows from the above equation that if the initial difference frequency Ω is less than $Kf(0)$ where $f(0)$ is the gain of the filter at d.c. a steady state corresponding to $\frac{d\phi}{dt} = 0$ will be reached and the phase error at steady state will be given by $\phi = \sin^{-1}[\Omega/Kf(0)]$. The other important parameters of a carrier-lock device are the locking range and the noise bandwidth which are determined by the filter function $f(p)$ and the gain constant K.

The carrier lock technique is obviously applicable in reception of only such modulated waves as contain a carrier. If, however, the carrier component is absent one has to rely on the sideband components for locking the receiver. It should be clear that a phase reference is available from the side-band components only in DSB (Double sideband) systems. A reference is absent in single sideband waves. In double sideband systems even if the carrier is suppressed a phase reference can be obtained from a study of the relations between the sideband components corresponding to a given modulating signal.

Sideband Lock : The receiver utilising 'sideband lock' technique (Costas, J. P., 1956) is based on quite different principles from those usual in practice, the frequency and phase synchronisation for the demodulating carrier being established from the received sideband components only. In the receiver the carrier information is ignored even if it is received along with the sideband. The principles of frequency and phase synchronism will be discussed in detail here. An alternative scheme implementing sideband lock technique will also be suggested in this connection.

Derivation of Phase Information from the Sideband Components : For the moment we assume that the demodulating carrier frequency is equal to that of the modulated carrier, whether the received signal is a transmitted carrier type or not. In the reception of DSB like signals the information required, is the arithmetic mean position between the sideband components resulting from modulation by a particular modulating frequency. The phase of demodulating carrier should be aligned in such a way as to have this arithmetic mean position. This will ensure coherent addition of the demodulated products of the received sideband components, that is, the modulating frequency outputs resulting from the upper and lower sideband components will be in phase with each other and the added products will be a maximum. Any departure in phase of the demodulating carrier from the value dictated by the arithmetic mean position will reduce the output (which is the sum of the contribution due to the two sideband components) by a factor cos ϕ_e where ϕ_e is the phase error mentioned. It is possible to circumvent this difficulty if we instead use a voltage controlled oscillator (VCO) in the receiver where a voltage sensitive capacitor serves as a tuning parameter that controls the oscillation phase of the receiver oscillator. The voltage sensitive capacitor can be actuated by a d.c. controlling voltage in proportion to the phase error talking about. The problem thus is to derive a d.c. controlling voltage in proportion to the phase error ϕ_e.

It can be shown that the demodulated product of the received signal and the demodulating carrier gives the required controlling voltage when multiplied with the demodulated product obtained from the same input but the demodulating carrier in this case being in phase quadrature with the former. The d.c. voltage (slowly varying) thus obtained can be shown to be proportional to the products of the amplitudes of the two sideband (the upper and lower) components and also to $\sin(2\phi_e)$ where ϕ_e is the phase departure. It is thus seen that the controlling voltage is sinusoidally varying with the phase error ϕ_e and is evidently zero when there is no phase error i.e. $\phi_e = 0$. The magnitude of this controlling voltage depends on the amount of phase error ϕ_e and the sense depends on whether ϕ_e is positive or negative. It is, therefore, clear that this error voltage may well be used to vary the capacitance value offered by the voltage sensitive capacitor, thereby controlling the oscillation phase of the VCO. Of course, it must be ascertained that the phase change in oscillation of the VCO caused by the capacitance change does reduce in fact the discrepancy. This depends on the sense of the change of the capacitance values caused by a particular polarity of the controlling voltage which in turn depends on the sign of the error ϕ_e. It may so happen that to obtain the right condition the sense of the controlling voltage is to be reversed and in such a case the VCO automatically comes in lock (synchronism) if the initial phase discrepancy is not greater than $\pm\pi/4$ radians and synchronism is established. A change in position of the received sideband components causes the arithmetic mean position to have a new value that causes a phase error, ϕ'_e, between the position of the demodulating carrier and the new arithmetic mean position which in turn causes the error voltage to have a new value that again nullifies or minimises the phase error. A little thought will show that such a system in any case cannot be a perfectly coherent device; there will always be a certain amount of phase discrepancy, however small. Under perfectly coherent conditions ϕ_e is equal to zero and the error voltage or the controlling voltage vanishes; in other words, there is no controlling voltage to control the phase of the demodulating carrier under perfectly synchronised condition. Therefore, it seems reasonable to call such a system to be a semicoherent device. The block schematic diagram for a phase locking arrangement following a SBPL technique is shown in fig. 3.

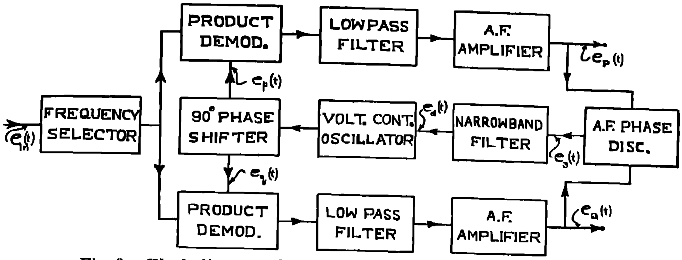

Fig. 3. Block diagram of a sideband phase locked reception system

Derivation of Frequency Information from the sideband components : In our previous discussions we devoted our attention to phase coherence with the assumption that frequency coherence already exists. It is of interest to see how frequency synchronism can also be achieved from the information derived purely from the sideband components. We need some controlling voltage which is proportional to the initial frequency discrepancy between the modulated carrier and the frequency of oscillation of the VCO. It is easy to see that the required controlling voltage is obtained from a frequency discriminator, an ideal multiplicative one, if we use the P and Q channel outputs and their time derivatives. It can be shown that the difference of the products $e_P(t)\,\dfrac{d}{dt}[e\ (t)]$ and $e_Q(t)\,\dfrac{d}{dt}\,[e_P(t)]$ gives a slowly varying d.c. voltage which is proportional to the frequency discrepancy Ω and may be used to bring down this discrepancy (Refer to fig. 4).

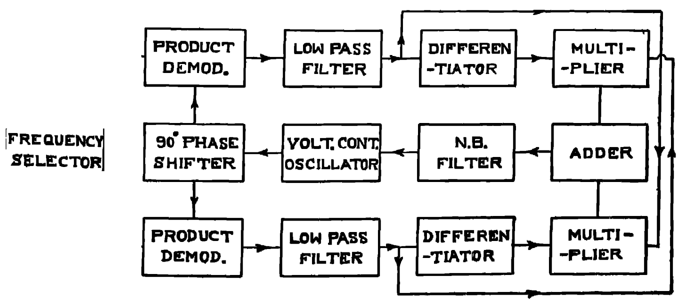

Fig. 4. Block diagram for establishing frequency synchronism from the sideband components

Sideband-Lock with DSB Signal: We shall now derive expressions for the controlling d.c. voltage and the inphase and quadrature channel outputs of such a receiving system when the input to the receiver is a DSB signal obtained by modulating a carrier by a single tone modulation. The input to the receiver (refer to fig. 2) in such a case may be written as

$$e_i(t) = A_+ \cos[(\omega_0+\omega_s)t+\phi_+]+A_- \cos[(\omega_0-\omega_s)t+\phi_-] \tag{2}$$

where A_+ and A_- are the received amplitudes, $\omega_0+\omega_s$ and $\omega_0-\omega_s$ are the frequencies, ϕ_+ and ϕ_- are the received phases of the upper and lower sideband components respectively resulting from a carrier $a\cos(\omega_0 t)$, and modulating signal $b\cos(\omega_s t)$.

The received signal is fed to two product demodulators and gets multiplied by two demodulating carriers $e_p = B\cos(\omega_0 t+\psi_0)$ and $e_q = B\sin(\omega_0 t+\psi_0)$ derived from the VCO but in phase quadrature with each other, ψ_0 being the static phase of the oscillator in the receiver.

The modulating frequency components from the output of the product demodulators are accepted by means of an a.f. filter and fed to an audio frequency phase detector. The inputs $e_1(t)$ and $e_2(t)$ to the a f. phase detector may be written to be

$$e_1(t) = \text{D.C.}\ \frac{B}{2}[A_+\cos(\omega_s t+\phi_+-\psi_0)+A_-\cos(\omega_s t-\phi_-+\psi_0)] \quad \ldots \quad (3)$$

$$e_2(t) = \text{D.C.}\ \frac{B}{2}[A_+\sin(\omega_s t+\phi_+-\psi_0)-A_-\sin(\omega_s t-\phi_-+\psi_0)] \quad \ldots \quad (4)$$

where C takes account of the constant of the product demodulator and low pass filter. D stands for the gain of the a f. amplifier stage.

The output from the a.f. phase detector, an ideal multiplicative one, may be written to be

$$e_3(t) = \frac{D^2C^2B^2}{8}\cdot k\,[A_+^2\sin 2(\omega_s t+\phi_+-\psi_0)-2A_+A_-\sin(\phi_++\phi_--2\psi_0)$$
$$-A_-^2\sin 2(\omega_s t-\phi_--\psi_0)].$$

where k stands for the constant of the a.f. phase detector.

The filter intervening the a.f. phase detector and VCO attenuates the modulating frequency components and components having higher frequencies and allows only the d.c. and slowly varying components to pass through it. The filter output may be written to be

$$e_d(t) = Kf(p)A_+A_-\sin(\phi_++\phi_--2\psi_0) \quad \ldots \quad (6)$$

where $f(p)$ stands for the filter transfer function and $K = \dfrac{D^2C^2B^2k}{4}$ takes account of all the constants including the amplitude of oscillation (B).

When synchronism is established the error voltage vanishes and the condition for coherence may be written to be

$$\phi_{+}+\phi_{-} = 2\psi_0 \quad \text{or,} \quad \psi_0 = \frac{\phi_{+}+\phi_{-}}{2} \qquad \ldots \quad (7)$$

i e., the phase of the local carrier equals the arithmetic mean of the phase dictated by the received sideband components. A departure in phase by an amount ϕ_e i e. $\psi_0 = \frac{\phi_{+}+\phi_{-}}{2} \pm \phi_e$ will cause an error voltage of magnitude $e_r = \pm Kf(p) A_{+}A_{-} \sin 2\phi_e$ to appear. Under perfectly locked condition $\phi_e = 0$ and $e_r = 0$ The outputs $e_1(t)$ and $e_2(t)$ change to $e_Q(t)$ and $e_P(t)$ and may be given by

$$e_P(t) = G(A_{+}+A_{-}) \cos(\omega_s t + \phi_{+} - \psi_0) \qquad \ldots \quad (8)$$

$$= G(A_{+}+A_{-}) \cos\left(\omega_s t + \frac{\phi_{+}-\phi_{-}}{2}\right) \qquad (8a)$$

$$e_Q(t) = G(A_{+}-A_{-}) \sin(\omega_s t + \phi_{+} - \psi_0) \qquad . \quad (9)$$

$$= G(A_{+}-A_{-}) \sin\left(\omega_s t + \frac{\phi_{+}-\phi_{-}}{2}\right) \qquad (9a)$$

where G has been substituted for DCB/2.

So we see that under locked condition the output $e_P(t)$ is a maximum and is proportional to the sum of the amplitudes of the received sideband components. On the other hand, the output $e_Q(t)$ is a minimum and is proportional to the difference of the amplitudes of the received sideband components which is evidently zero when the received amplitudes of the two sideband components are equal. It is interesting to see that the demodulating carrier in the product demodulator which ultimately gives the output is in phase with the modulated carrier and the demodulated products of this channel may accordingly be called in phase channel or simply P channel output. The demodulating carrier in the other product demodulator is in phase quadrature and the demodulated products of this channel may accordingly be called quadrature channel or simply Q channel output.

It has already been pointed out that perfect coherence can never be achieved in such a closed loop error actuated system and there will always be a discrepancy ϕ_e in phase, however small. In such a case ψ_0 may be written to be $\psi_0 = \frac{\phi_{+}+\phi_{-}}{2} \pm\phi_e$ and the P and Q channel outputs may be written as

$$e'_P(t) = G\Big[(A_{+}+A_{-}) \cos\left(\omega_s t + \frac{\phi_{+}-\phi_{-}}{2}\right) \cos\phi_e$$

$$\pm(A_{+}-A_{-}) \sin\left(\omega_s t + \frac{\phi_{+}-\phi_{-}}{2}\right) \sin\phi_e\Big] \qquad (10)$$

$$\simeq G(A_+ + A_-)\cdot\cos\phi_e\cdot\cos\left(\omega_s t+\frac{\phi_+-\phi_-}{2}\right) \tag{10a}$$

$$e'_Q(t) = G\Big[(A_+ - A_-)\sin\left(\omega_s + \frac{\phi_+-\phi_-}{2}\right)\cos\phi_e$$

$$\mp(A_+ + A_-)\cos\left(\omega_s t+\frac{\phi_+-\phi_-}{2}\right)\cdot\sin\phi_e\Big] \tag{11}$$

$$\simeq \quad \mp G\cdot(A_+ + A_-)\sin\phi_e\cdot\cos\left(\omega_s t+\frac{\phi_+-\phi_-}{2}\right) \tag{11a}$$

In the following we shall treat the case considered above when the input to the receiver is a DSB signal resulting from multitone modulation.

When the modulating signal is complex, consisting of a large number of tones, the input voltage may be written as

$$e_i(t) = R_e \Sigma A_u \exp j[(\omega_0+u)t+\phi_u] \tag{12}$$

The output of the product detector can then be represented by

$$e_d(t) = \Sigma\ \Sigma\ A_u A_v \sin\ [(u+v)t+\phi_u+\phi_v-2\psi_0] \tag{13}$$

where A_u and A_v are the amplitudes corresponding to frequency components u and v and ψ_0 is the phase of the oscillator voltage. This expression is the same as that results from squaring the input and the L.O. voltage and feeding these to a phase detector.

If the oscillator contains phase modulation (which is indeed the case when it is tracking), the L.O. voltage can be written as

$$e_{L\cdot O}(t) = B_W \exp j[(\omega_0+w)t+\psi_W] \quad \ldots \tag{14}$$

The output of the a.f. phase detector in such a case is

$$e_d(t) = \sum_u \sum_v \sum_w \sum_x A_u A_v B_w B_x \sin[(u+v)t-(\omega+x)t+\phi_u+\phi_v-(\psi_w+\psi_x)] \quad .. \tag{15}$$

Low frequency components of the output in a band B_F correspond to the condition

$$|u+v-(\omega+x)| \leqslant B_F$$

It can easily be shown that if the signal input to the receiver and the signal from the local oscillator are squared and mixed one gets the same phase equation as in DSB phase lock. Recognition of this equivalence enables one to simplify the locking procedure in coherent reception of VSB signals. (The normal technique is to establish the phase coherence with the second harmonic component of the local oscillator). When synchronism is established the P channel in the DSB receiver gives the modulating signal output.

It should be observed that a sideband component ω_0+u say, is effective in producing an error voltage mixing with the sideband component ω_0+v at the output of the filter only when $u = v$. If $u \neq v$ an a-c voltage too appears at the output of the a.f. phase detector of frequency $(u{\sim}v)$ which is however attenuated by the filter.

Domination of the stronger components in phase locking : The expression for the loop phase equation (Refer eq. 13) shows that the components of signal contribute to the locking voltage in proportion to their powers. Consequently the final equillibrium phases is determined by the stronger components. It follows that if the values of the mean carrier phases demanded by the sideband components in the different parts of the band vary between wide limits, it is only the strong components for which the phase adjustment is correct

A problem that arises in the case of such two loop systems is that of relative phasing between the output in the two bands involved. (In speech modulation this problem is not important). It is easy to see that for a keyed waveform if the relative phases between the different parts of the spectrum are not maintained, the waveform may be distorted without recognition.

Bandwidth limitation in the locking loop . It is known that the spectral distribution of the modulated signal is in general non-uniform and the total power in part of the spectrum defined by $B/2 < |f-f_0|$ is a small fraction of the power contained in the region defined by $B/2 > |f-f_0|$, where f_0 is the centre frequency and $2B$ is the total r f. bandwidth Use can be made of this information for improving the signal to noise ratio in the locking loop by restricting the bandwidth of the video or audio filter. This restriction will cause a reduction of the signal power but this is outweighed by the improvement in SNR obtainable, which is given by

$$F = \frac{\text{Signal power density in the reduced band}}{\text{Signal power density in the original band}}$$

Interference Rejection : The reception technique discussed with reference to DSB modulation may well be utilised to have a completely interference free demodulated product in case of severe interferences that smudge out the information bearing signal in any one of the two sidebands. The interfering signal may be characterised by

$$e_i(t) = I \cos[(\omega_0 \pm \omega_i)t+\phi_i] \qquad \cdots \quad (16)$$

according as the interfering signal falls within the upper $(\omega_0+\omega_i)$ or lower $(\omega_0-\omega_i)$ sideband. It may easily be seen that the P and Q channel outputs contain the contribution from the interfering component in addition to the information bearing components. The interfering component in the demodulated products can be annulled if the outputs $e_P(t)$ and $e_Q(t)$ are given phase shifts θ_P and θ_Q such that

$\theta_P \sim \theta_Q$ is equal to $\pi/2$ radians and the phase shifted outputs are added or subtracted according as the interfering signal is in the lower or upper sideband in the received modulated signal.

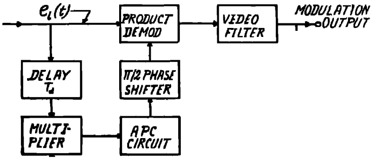

Fig 5. Block diagram of an alternative scheme suggested for sideband phase locked reception

An alternative scheme for sideband phase locking: In this scheme (Refer fig. 5) the incoming signal is fed to a product demodulator where it gets multiplied by the coherent demodulating carrier. The modulation component is accepted by means of a video filter from the products of the demodulator. The filter output gives the modulation or keyed information.

The locking loop consists of a bandpass filter, a multiplier and a CPL circuit. Use is made of a band pass filter to introduce a time delay to the incoming r.f. signal, the amount of delay being equal to that suffered by the modulation components in the video filter. The purpose of the CPL circuit followed by the multiplier is to derive the necessary demodulating carrier out of the products of the multiplier. A little thought will show that the expressions for the controlling d.c. voltage in this case are identical with the equations (6) and (13) for single tone and multitone DSB modulation. When the received signal $e_i(t) = \pm F(t) \sin(\omega_0 t + \phi_0)$ (i.e. Binary PSK modulation) the controlling voltage is given by $e_d(t) = KF^2(t - T_d) \sin 2(\phi_0 - \psi_0)$, where T_d is the time delay mentioned.

RECEPTION OF BINARY PSK, CONVENTIONAL AM, SSB AND NARROWBAND PM OR FM SIGNALS FOLLOWING A SBPL TECHNIQUE:

Binary PSK: In Binary PSK modulation the phase of the transmitted carrier is altered between two states (either 0 or π) in accordance with the transmitted message. A binary PSK signal may be generated by modulating the carrier in amplitude by a square wave alternating between the amplitudes $\pm a$. The carrier is suppressed as in DSB modulation.

The received signal in this case may be written to be

$$e_i(t) = \pm A \cos(\omega_0 t + \phi_0) \qquad \dots (17)$$

The demodulating carrier is brought in synchronism with the modulated one following a DSB sideband lock technique. When locking is ensured i.e.

$\psi_0 = \phi_0$ the P channel gives an output in accordance with the state of the modulating signal i.e. either $+A$ or $-A$.

Conventional AM : The received signal in this case may be written to be

$$e_i(t) = A_0 \cos(w_0 t+\phi_0)+\Sigma A_u \cos[(\omega_0+u)t+\phi_u]+\Sigma A_v \cos[(\omega_0-v)t+\phi_v] \qquad (18)$$

Since the carrier component is present, one should make use of the carrier for measurement of the frequency.

The use of an AFC circuit operating on the carrier component would provide an additional facility to maintain the required amount of frequency stability to the VCO. The P channel in this case gives the modulation frequency input.

SSB Modulation : The received sideband in this case may be characterised by either

$$e_i(t) = \Sigma A_u \cos[(\omega_0+u)t+\phi_u] \qquad \ldots (19)$$

or,

$$e_i(t) = \Sigma A_v \cos[(\omega_0-v)t+\phi_v] \qquad \ldots (19a)$$

The reception of this type of signal may be effected using a crystal controlled oscillator to ensure good frequency stability. In this case both P and Q channels contain the modulating signal components and are in phase quadrature. The outputs of both the channels are to be added up after appropriate phasing. The outputs $e_P(t)$ and $e_Q(t)$ are given phase shifts θ_P and θ_Q such that $\theta_P \sim \theta_Q = \pi/2$ over the audio band and the phase shifted products are added or subtracted to give the modulation output. It should be pointed out in this case that reception can only be effected for signals resulting from speech like modulation. In the case of speech it has been found that intelligibility is not seriously impaired, as long as the local oscillator is within 10 cycles of the correct frequency, even though there may be some loss in naturalness.

Narrowband PM or FM : The received signal in this case may be written to be

$$e_i(t) = A \cos(\omega_0 t+m \cos \omega_s t) \qquad \ldots (20)$$

$$\simeq A \cos \omega_0 t - mA \sin \omega_0 t \cdot \cos \omega_s t. \qquad \ldots (20a)$$

for small values of m.

We thus have, to a first approximation, a carrier $A \cos \omega_0 t$ and two sideband components, from a quadrature carrier, $mA \sin \omega_0 t \cdot \cos \omega_s t$.

The demodulating carrier in the P channel in this case is aligned with the arithmetic mean position dictated by the sideband components. The demodulated output is obtained from this channel. If the device be a purely sideband locked type, the Q channel output will be a minimum. In fact the demodulating carrier in the P channel is in phase quadrature with the incoming carrier and that of the Q channel is in phase with it.

V.S.B. Modulation : The spectrum of VSB is asymmetric and ordinarily the transmission at carrier frequency is a fraction of the maximum. A VSB wave $S(t)$ can therefore be resolved into two components, an inphase component $P(t)$ and a quadrature component $Q(t)$. Thus

$$S(t) = P(t) \cos(\omega_0 t+\phi_0)+Q(t) \sin(\omega_0 t+\phi_0) \tag{21}$$

$$= E(t) \cos[\omega_0 t+\phi_0+\phi(t)] \tag{21a}$$

where $E(t) = \sqrt{P^2(t)+Q^2(t)}$ is the envelope, $\phi(t) = \tan^{-1} \dfrac{Q(t)}{P(t)}$ is the phase and ω_0 is the angular frequency of the carrier.

In receiving a VSB modulation one has to find the average inphase component in the received signal. For the purpose the input is multiplied by itself and the second harmonic component is selected. This output is given by

$$S_2(t) = [P^2(t)-Q^2(t)] \cos 2(\omega_0 t+\phi_0)+P(t) \cdot Q(t) \sin 2(\omega_0 t-\phi_0) \tag{22}$$

A locally generated carrier $B \cos(\omega_0 t+\psi_0)$ is locked inphase with the second harmonic component by shifting the voltage by $\pi/4$ radians, squaring it and mixing the doubled output with $S_2(t)$. The phase of the oscillator is controlled by means of an average value of the voltage derived out of mixing. The controlling voltage can be written as

$$e_d(t) = [P^2(t)-Q^2(t)] \sin 2(\phi_0-\psi_0)+P(t) \cdot Q(t) \cdot \cos 2(\phi_0-\psi_0) \tag{23}$$

Now the average value of the product $P(t) \cdot Q(t)$ is zero and the average value of $Q^2(t)$ is small compared to that of $P^2(t)$. The average value of the voltage $e_d(t)$ is then approximately equal to

$$e_d = KP^2(t) \cdot \sin 2(\phi_0-\psi_0). \quad \ldots \tag{23a}$$

which is identical with that obtained in eq. (6).

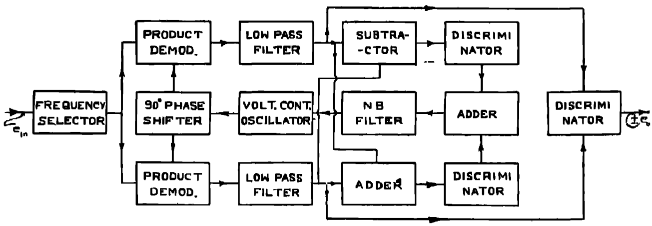

Fig. 6. Block diagram for the reception of Narrowband FSK signals

Narrowband FSK : Coherent reception of narrowband FSK signals is possible following SBPL technique. In this case two low frequency discriminators are to be incorporated in the upper sideband and lower sideband channels. The outputs of these discriminators may be used to maintain the frequency coherence of the local oscillator (Refer fig. 6).

EFFECTS OF NOISE AND INTERFERENCE ON THE CAPTURE CAPABILITY OF THE PHASE LOCKED LOOP (PLL):

The input signal to a receiver in the presence of an additive noise may be written as

$$e_i(t) = Re \exp(j\omega_0 t)[A_0 \exp(j\phi_0) + \Sigma A_k \exp j(\omega_k t + \phi_k) + \Sigma n_l \exp j(\omega_l t + \beta_l)] \quad \ldots \quad (24)$$

where Re stands for the real part of, A_0, ω_0 and ϕ_0 denote the amplitude, frequency and phase for the carrier component respectively, A_k and ϕ_k are the amplitude and phase of the sideband component at an angular frequency $\omega_0 + \omega_k$ and n_l, ω_l and β_l are the values of the corresponding quantities for the narrowband noise process.

With reference to DSB modulation reception (Ref. fig. 3) the injected demodulating carriers may be represented by

$$e_p(t) = Re\,[\exp j(\omega_0 t + \psi_0) + \exp -j(\omega_0 t + \psi_0)] \text{ in the } P \text{ channel} \quad \ldots \quad (25)$$

$$e_q(t) = Im\,[\exp j(\omega_0 t + \psi_0) - \exp -j(\omega_0 t + \psi_0)] \text{ in the } Q \text{ channel} \quad \ldots \quad (26)$$

The low frequency outputs of the P and Q channel demodulators may be written to be

$$e_P(t) = Re[A_0 \exp j(\phi_0 - \psi_0) + \Sigma A_k \exp j(\omega_k t + \phi_k - \psi_0) + \Sigma n_l \exp j(\omega_l t + \beta_l - \psi_0)] \quad \ldots \quad (27)$$

$$e_Q(t) = Im[A_0 \exp j(\phi_0 - \psi_0) + \Sigma A_k \exp j(\omega_k t + \phi_k - \psi_0) + \Sigma n_l \exp j(\omega_l t + \beta_l - \psi_0)] \quad \ldots \quad (28)$$

The phase detector output $e_d(t) = e_P(t) \cdot e_Q(t)$ is given by

$$\begin{aligned} e_d(t) = k[&A_0^2/2 \sin 2(\phi_0 - \psi_0) + A_0 \Sigma A_k \sin(\omega_k t + \phi_k + \phi_0 - 2\psi_0) \\ &+ \Sigma\ \Sigma A_u A_v \sin(\quad t + \phi_u + \phi_v - 2\psi_0) \\ &+ \Sigma\ \Sigma A_k n_l \sin(\overline{\omega_k + \omega_l} t + \phi_k + \beta_l - 2\psi_0) \\ &+ A_0 \Sigma n_l \sin(\omega_l t + \beta_l + \phi_0 - 2\psi_0) + \Sigma\ \Sigma\ n_w n_x \sin(\overline{\omega + x} t + \beta_w + \beta_x - 2\psi_0)] \end{aligned} \quad \ldots \quad (29)$$

Phase Equations : Let us now examine the phase equations in presence of an additive noise in CPL and SBPL cases.

Case I. Carrier Phase lock :

Let the received signal be given by

$$e_i(t) = A_0 \cos(\omega_0 t + \phi_0) + n_c \cos(\omega_0 t + \phi_0) + n_q \sin(\omega_0 t + \phi_0) \quad \ldots \quad (30)$$

where n_c and n_q are the noise components in phase and in quadrature with the carrier $A_0 \cos(\omega_0 t+\phi_0)$, white gaussian, independent variables with the highest frequency Δf. Further $\overline{n_c^2} = \overline{n_q^2} = 2n_0^2\Delta f$ ... (31)

where n_0^2 is the input noise power density expressed in watts per cycle. The output of the product detector will be (see eq. 29)

$$e_d(t) = k[A_0 \sin\phi + n_c \sin\phi + n_q \cos\phi] \quad \text{... (32)}$$

If the phase error ϕ is small,

$$e_d(t) = k[(A_0+n_c)\phi + n_q] \quad \text{... (33)}$$

If this output be used to control the phase of the oscillator, then

$$\phi(t) = -k_1 f_1(p) e_d(t). \quad \text{... (34)}$$

one derives that $\phi(t) = \dfrac{k_1 f_1(p) \cdot n_q}{1+(n_c+A_0)k_1 f_1(p)}$ ... (35)

If $n_c\phi$ is small compared to A_0, $\phi \simeq \dfrac{n_q}{A_0}$ ... (36)

If, on the other hand, the phase detector output controls the frequency of the oscillator, then

$$\begin{aligned}\frac{d\phi}{dt} &= \Omega - k_1 f_1(p) e_d(t) \\ &= \Omega - k_1 f_1(p) \cdot [(A_0+n_c)\sin\phi + n_q \cos\phi] \\ &\simeq \Omega - k_1 f_1(p)[(A_0+n_c)\phi + n_q]\end{aligned} \quad (37)$$

If $n_c\phi$ is small,

$$\phi = \frac{\Omega - k_1 f_1(p) n_q}{p + k_1 A_0 f_1(p)}. \quad (38)$$

Case II. Sideband phase lock :

Let us consider single tone modulation and restrict attention to noise components around the sidebands. The received signal in this case may be given by

$$\begin{aligned}e_i(t) = [&A_+ \cos\{(\omega_0+\omega_s)t+\phi_+\} + A_- \cos\{(\omega_0-\omega_s)t+\phi_-\} \\ &+ n_{c_+} \cos\{(\omega_0+\omega_s)t+\phi_+\} + n_{c_-} \cos\{(\omega_0-\omega_s)t+\phi_-\} \\ &+ n_{q_+} \sin\{(\omega_0+\omega_s)t+\phi_+\} + n_{q_-} \sin\{(\omega_0-\omega_s)t+\phi_-\}]\end{aligned} \quad \text{... (39)}$$

where all the noise voltages are white, gaussian, independent and with the highest frequency Δf cycles. Further $\overline{n^2_{c+}} = \overline{n^2_{c-}} = \overline{n^2_{q+}} = \overline{n^2_{q-}} = 2n^2_0\Delta f$ ... (40)

If the phase detector output given by eq. (29) is used to control the frequency of the oscillator, the loop phase equation can be written as

$$e_d(t) = \frac{1}{k_2 f_2(p)} \cdot \frac{d\phi}{dt}$$

$$= [A_+ A_- \sin(\phi_+ + \phi_- - 2\psi_0) + (A_+ n_{q-} + A_- n_{q+}) \cos 2\phi$$

$$+ (n_{c+} n_{c-} + n_{q+} n_{q-}) \sin 2\phi + (n_{c+} n_{q-} - n_{c-} n_{q+}) \cos 2\phi] \quad \ldots \quad (41)$$

If ϕ is small one may write to a first order, assuming $A_+ = A_- = A$ say,

$$e_d(t) = k \left[\frac{A^2}{2} \phi + \frac{A}{2} (n_{q+} + n_{q-}) + n_{c+} n_{q-} - n_{c-} n_{q+} \right] \quad \ldots \quad (41a)$$

The first term is the desired phase control voltage and the remaining terms are the undesired noise voltages. The noise power appearing at the output of the filter can be found thus

$$\overline{e^2_d} = k^2 [A^2/4 \cdot (\overline{n^2_{q+}} + \overline{n^2_{q-}}) + \overline{n^2_{c+} n^2_{q-}} + \overline{n^2_{c-} n^2_{q+}} + 2(\overline{n_{c+} n_{q-} - n_{c-} n_{q+}})] \quad \ldots \quad (42)$$

Noise and interference : It is clear from the expression for the phase locking voltage that in CPL technique the noise and interference components located round the carrier only are responsible in producing the control voltage in the loop. The effects of these components will be to increase the instantaneous phase error between the incoming carrier and the demodulating carrier in the receiver.

When SBPL only is used the components of the noise and interference located round the carrier have no adverse effect as far as the controlling loop is concerned; those affecting the loop are the components which are located round the sideband components arising from the modulation. It will be observed that in this case a part of the output will be produced due to intermodulation between the noise components in the audio band. It will be seen that the appropriate error voltage (proportional to the phase difference between the two carriers) for CPL can be obtained from the quadrature output. Therefore if CPL alone is desired this component after appropriate filtering should only be used for the purpose of locking. It should be mentioned that sometimes the phase of the received carrier may depart substantially from the position indicated by the sidebands. Since the modulation is present essentially in the sidebands the demodulated voltage using the received carrier as the reference is likely to be smaller than that is obtainable by using the sideband reference. In such case it is desirable to use the CPL voltage for adjusting the frequency of the local carrier and the SBPL voltage for adjusting its phase.

The loop phase equation for a loop employing both CPL and SBPL can be written as

$$\frac{d\phi}{dt} = \Omega - k_1 f_1(p)\Big[\frac{A_0^2}{2}\sin 2(\phi_0-\psi_0)+A_0\Sigma A_k \sin(\omega_k t+\phi_k+\phi_0-2\psi_0)$$
$$+\Sigma\Sigma A_u A_v \sin(ut+vt+\phi_u+\phi_v-2\psi_0)\Big]$$
$$-k_2 f_2(p)[A_0 \sin(\phi_0-\psi_0)+\Sigma A_k \sin(\omega_k t+\phi_k-2\psi_0)] \quad \ldots \quad (43)$$

where $k_1 f_1(p)$ and $k_2 f_2(p)$ are the gains associated with the control loops.

Deterioration of the SNR due to noise present in the PLL : Any noise present in the input will cause a perturbation of the phase of the VCO. The effect of the phase noise will cause a reduction of the SNR at the output from the corresponding value at the input due to the intermodulation.

Assuming that the phase perturbation caused by the noise is given by ϕ_{on} it is easy to see that the outputs of the P and Q channels are respectively

$$e_P(t) = [(A_+ + A_-)\cos(\phi_e+\phi_{on})\cos(\omega_s t+\psi)$$
$$+(A_+ - A_-)\sin(\phi_e+\phi_{on})\sin(\omega_s t+\psi)$$
$$+n_c \sin(\phi_e+\phi_{on})+n_q\cos(\phi_e+\phi_{on})] \quad \ldots (44)$$

and

$$e_Q(t) = [(A_+ - A_-)\cos(\phi_e+\phi_{on})\sin(\omega_s t+\psi)$$
$$-(A_+ + A_-)\sin(\phi_e+\phi_{on})\cos(\omega_s t+\psi)$$
$$+n_c\cos(\phi_e+\phi_{on})+n_q\sin(\phi_e+\phi_{on})] \quad \ldots \quad (45)$$

This shows that the modulation output in the P channel decreases whereas the noise output remains constant, that is to say that the output SNR is smaller than that could have been for perfect locking.

EXPERIMENTAL SET-UP, RESULTS AND DISCUSSIONS

A transistorized version of a laboratory model of a receiving system implementing SBPL technique has been constructed and tested. The objective of

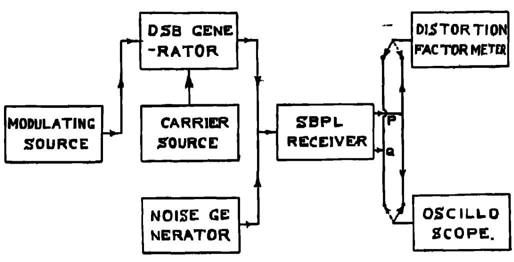

Fig. 7. Block diagram for the experimental arrangements to study the applicability of CPL and SBPL techniques in DSB modulation reception.

the experimental work was to provide experimental verification of some of the conclusions arrived at the previous sections and also to demonstrate the basic principle of operation of the system.

We shall describe here the experimental arrangements for studying CPL and SBPL techniques in reception of different modulations and diversity combination and the results obtained experimentally. A discussion of the experimental results, and some remarks on the design fabrications and performances of such a receiving system, are also given. Studies have been made of the characteristics of the receivers employing sideband phase lock technique in regard to reception of DSB- AM, NBPM and also Binary PSK signals. For these studies complete receivers embodying the techniques mentioned have been constructed. The test equipment constructed include DSB generator, NBPM generator and random fading generator.

Circuit and Experimental Arrangements : We shall first describe the auxiliary circuits necessary in the experimental arrangements for studying the applicability of CPL and SBPL techniques in reception of different types of modulated signal and diversity combination. The most important components of any locked receiver are evidently the phase and frequency control circuit incorporating the phase and frequency detector, voltage controlled phase shifters and voltage controlled oscillators. In what follows (the descriptions, performances, circuit diagrams of the units actually used together with typical characteristics obtained experimentally of) *some of the circuits will be briefly discussed.*

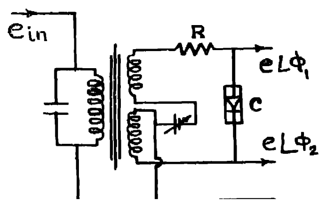

Fig. 8. Circuit diagram of a voltage controlled phase shifter.

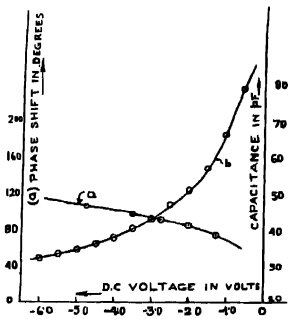

Fig. 9. Figure shows a typical phase voltage characteristic of the voltage controlled phase shifter.

Voltage Controlled Phase Shifter—In a voltage controlled phase shifter the amount of phase shifts obtainable is a function of some d.c voltage. A voltage sensitive diode is used in the phase shifter. The magnitude of the capacitance thrown by such a diode depends on the magnitude of the d.c. potential across its terminals. fig. (8) shows the circuit diagram for such a phase shifter and fig. (9) is a typical plot showing the variation of phase shifts with d.c. voltage using a diode type V No. 33 in the phase shift network.

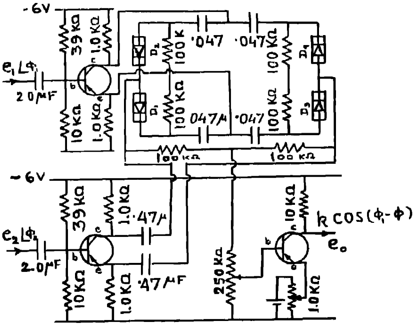

Fig. 10. Circuit diagram of an a.f. phase discriminator in the receiver used to derive the necessary controlling voltage for the VCO's.

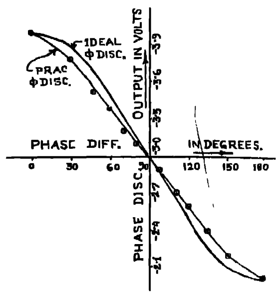

Fig. 11. Figure shows the variation of the d.c. output voltage with phase difference between the two inputs of the audio frequency phase discriminator

A.F. Phase Discriminator—The circuit diagram for an audio frequency phase discriminator is shown in fig. (10). When the two inputs e_1 and e_2 are of the same frequency the output for such a system can be shown to be very nearly equal to $KE_2 \cos \phi$, where E_2 is the amplitude of e_2, ϕ stands for the difference in

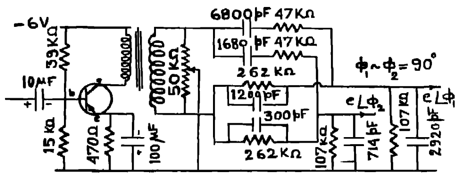

Fig. 12. Circuit diagram of an audio frequency wideband 90° phase shifter.

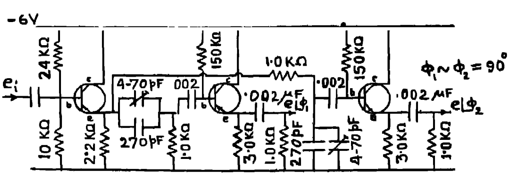

Fig. 13. Circuit diagram of a radio frequency 90° phase shifter.

phase between e_1 and e_2 and K is a constant for the discriminator under condition that $E_2 << E_1$, the amplitude of e_1. fig. (11) shows a set of typical characteristics obtained in an actual experiment which depict the variations of output voltage (d c) with difference of phase (ϕ), for different amplitude ratios (i e. E_2/E_1).

Experimental results and discussions : For any measurement on receiving devices making use of phase lock techniques it must first be ensured that the system has adequate dynamic range both in the locking and demodulating circuits and also satisfactory locking capabilities both in respect of locking range and locking time. One has here to take into consideration the extent and rate of variation of the received input signal and design the control loop accordingly. This aspect of the problem will be discussed later.

Assuming that the loop has been properly designed, one has to test whether the reception technique employed is capable in practice of giving a satisfactory quality of the output signal.

In the receiver employing SBPL technique, measurements have been made with regard to characteristics of reception for DSBAM, NBPM and Binary PSK signals.

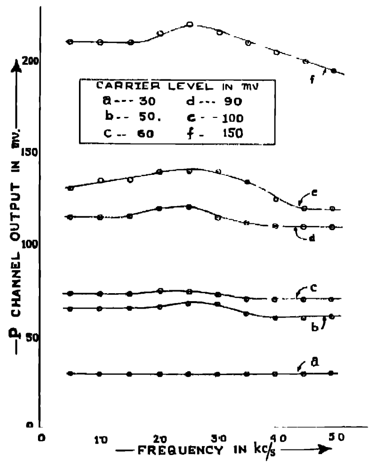

Fig. 14. Figure shows the variation of P channel output with modulating frequency and levels of the modulated input to the receiver for a single tone DSB modulated signal.

Experimental results with single tone modulating signal are presented in fig.(14) where the variations of the output in the P channel with frequency are shown for different levels of the input signal.

Measurements of the output signal and noise components have also been taken in presence of an additive noise at the input to the receiver. Fig 15 shows the variation of the output signal and noise components with different levels of the input noise.

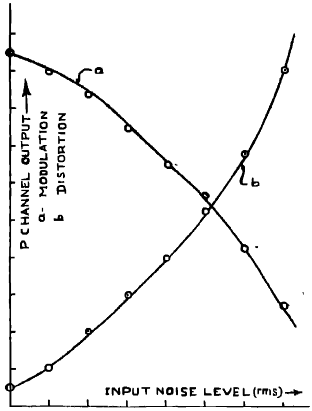

Fig. 15 Figure shows the variations of the output signal and noise levels with input noise level while the input signal level is held constant and that due to a single tone DSB modulation.

The effect of phase jitter caused by the transmission media to the input carrier was also studied in the experimental scheme. The phase jittering was simulated by phase modulating a r.f. carrier (amplitude modulation was balanced out) at a slow rate (of the order of a few tens of cycles), and using this phase modulated signal as the modulated carrier in DSB signal generation. The effect observed in the demodulated product was an amplitude modulation at a rate equal to the frequency of the carrier jitter.

The effect of a CW interfering carrier was also studied. In the experimental scheme using a 1.0 Kc/s modulating signal in DSB modulation, a sudden increase in the channel output was observed when the interfering signal was detuned from the incoming carrier by 1.0 Kc/s on either side of the latter. The fact is in agreement with the analysis in section 4.2.

In receiving Binary PSK signal it was observed that the *P* channel output gives the keyed information and the *Q* channel output remains substantially low so long the system is in lock. From the studies of additive noise in this case it has been confirmed that the reception of a binary PSK signal can be successfully carried out with the *DSB* receiver if the input SNR is adequate.

In reception of NBPM or NBFM signal it is experienced that the incoming carrier is partly being injected into the demodulating carrier and tries to get it locked with the latter even if the sideband lock voltage is reduced to zero, which causes the ratio of the outputs in the P and Q channels to be not as good as in DSB modulation reception. It seems advisable to use APC techniques for reception of such signals.

DISCUSSIONS

The following observations with regard to DSB signal reception may be made from the above measurements :

(i) Input r.f. amplitude must lie between certain limits for proper operation. If the input amplitude is too low the system does not lock If, on the other hand, the input is too high the system also loses lock. This is possibly due to the fact that (since the closed loop bandwidth is larger for larger input amplitude) loop bandwidth at large value of the input amplitude is such as to permit second harmonic of the input modulation to circulate within loop and cause disruption.

(ii) Adequate locking is established if the modulating frequency for the DSB signal lies between 100 cps to 10 Kcps. Both the limits can be extended by controlling the loop parameters (gain etc.).

(iii) When the system is in lock, the output of the inphase channel will be a maximum and that of the quadrature channel will be a minimum. The actual ratio of these two outputs depends on the following :

(a) Stability of the local oscillator frequency.

(b) Phasing in the demodulating carriers.

(c) Phase shift characteristics of the P and Q channels.

(d) Initial difference frequency between the incoming carrier and the local carrier.

(e) Phase and amplitude relationship between the two sidebands.

Under conditions of correct adjustment and if $(A_+ = A_-)$, $(\Omega = 0)$, $(\psi_{p,q} = \pi/2,\ \phi_0 - \psi_0 = 0)$ this ratio is found experimentally to be about 0.05.

It is extremely important to ensure a fair amount of stability of the oscillator frequency when pseudo-static locking as necessary in DSB is to be established, and to have a better stability in the feedback loop. For DSB reception purposes it is expected that a crystal oscillator with frequency controllable within 400 c/s from its centre frequency will serve to remove the instability experienced in the system.

Considerations in design of the phase locking loop : In all coherent reception techniques the greatest single requirement is that the system remains in lock in presence of noises and interferences with the carrier and modulation power available. The effective input power contributed by the carrier and modulation components will in general vary between fairly wide limits. One cause of this variation is the phenomenon of fading in r.f. circuits. Further, the modulation process may be such that the input power varies considerably. For example in linear systems like DSB and SSB the total power is obviously a function of the strength of instantaneous modulation, which for speech-like signals fluctuates by amounts exceeding 30 to 40 dbs. In amplitude keyed signals or phase or frequency keyed signals also the instantaneous amplitude of the received carrier varies with time.

Considering now the noise and interference powers one notes that these have fairly high peak factors (peak/rms). In the case of fluctuation noise, for example, the peak factor is as high as 4. It is clear then that the r.m.s. values of these interferences do not give an useful and correct estimate of their ability to disturb the circuit.

The choice of the system bandwidth is obviously a matter of compromise. Any design of locked receiver must take these factors into consideration to ensure that the circuit remains usable under the wide range of operating condition likely to prevail at one time or another. There will be two main effects of the noises and interferences present in the system. The first is that the system may lose lock and the second is that although the system remains in lock the SNR at the output is so very low that the system is useless. If the bandwidth of the locking system is small compared with the bandwidth of the modulating frequencies the second phenomenon may occur earlier as the input noise to the system increases. If the badwidths are of the same order both may occur simultaneously.

It is certainly desirable to keep the noise bandwidth of the control loop as small as possible in order that the noises and interferences present along with the signal do not cause system to lose lock with the desired signal phase. Further large loop bandwidth will increase the probability of faulty locking of the local carrier with the incoming carrier and one sideband component when a fraction of the carrier is also present along with the sidebands. The demodulating carrier frequency may also vary as the modulation varies. A narrower system bandwidth reduces the ability of the receiver to follow such changes. It must however be ensured that the system is able to follow the perturbations of the signal phase caused by the fluctuations of the medium. These two requirements define the upper and lower limits of the closed loop noise bandwidth of the control circuit. As mentioned earlier a further complication arises out of the fact that the signal amplitude varies with time and the closed loop noise bandwidth being dependent on loop gain also varies with this amplitude. It is, therefore, preferable to select only such lowpass filters as do not condition a large change in noise-bandwidth for changes in loop gain and also to incorporate appropriate automatic gain control circuits. Besides low audio frequency components below 300 c/s should suffer very large attenuation in the AF circuit feeding the phase discriminator in the sideband lock circuit. For otherwise these may circulate in the loop and cause disruption of the control loop.

CONCLUDING REMARKS

The results of this paper show that a coherent receiver utilising the information contained in the carrier and sideband components of the transmitted signal for locking the demodulating wave in the receiver is realisable in practice without undue complications provided the signal to noise ratio is adequate. The locking techniques studied have been classified as (a) carrier lock, (b) sideband lock techniques; the former is appropriate for demodulation of CW type signals and the latter for AM like signals. It should be observed that in the carrier lock case initial phase discrepancy between the incoming carrier and the demodulating one should not be greater than $\pi/2$ rad. i.e., $|\phi_0 - \psi_0| \leqslant \pi/2$ whereas in the sideband lock case this discrepancy is to be within $\pi/4$ rad. i.e., $|(\phi_+ + \phi_-)/2 - \psi_0| \leqslant \pi/4$.

It should be emphasised that an efficient and satisfactory locking system should take note of all the apriori information (viz., frequency, phase and time of occurrence) available for the received signal. The nature of medium variations and the knowledge of the waveform of the transmitted message (in reception of keyed signals) should also be utilised to have a satisfactory coherent reception

It should be obvious that techniques of locking can be utilised only if the average perturbation of the parameters over the modulation period are small. However if the power capacity of the transmitter be small and hence the SNR is low, message intervals will have to be made long in order that energy per symbol may exceed a minimum permissible value necessary for satisfactory reception. It may so happen in such a situation that information about frequency and phase desired to be estimated and stored may undergo large fluctuations over the message period. It is clear that any attempt at coherence worth the name is foredoomed if the correlation time of fluctuation is comparable with the individual message period. In such a case it is advisable to adopt what is called differentially coherent reception, where information embedded in the signal in the immediate past interval only are used as the reference for detection of a currently received signal. The reference is obviously a non-ideal one but that is possibly the best available in the circumstances.

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